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Co-Simulation of RFIC with Bondwire Antenna via Retarded PEEC Method

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Abstract—We present an antenna modeling method based on partial element equivalent circuit (PEEC) theory. The antenna is modeled as an equivalent circuit of lumped circuit elements, which enables circuit co-simulation between the antenna and circuits in both time and frequency domains. Antenna radiation is captured as an equivalent radiation resistor. For verification, a 2.4-GHz transmitter with bondwire antenna was implemented in a standard digital 0.35-μm CMOS technology. Measurement shows good agreement with the proposed model. This model can be applied to any other on-chip electromagnetic structures.

I. INTRODUCTION

In recent years, short-range communications have been embedded in a huge number of consumer electronic applications with compact physical sizes and efficient costs. Among these systems, integrated antenna is one of the critical elements to obtain good range and stable throughput. Normally, an integrated antenna is implemented in the form of an on-chip planar antenna [1]. However, this approach occupies large valuable silicon area and the resulted antenna efficiency is extremely low. Application of bondwire as antenna was reported in [2], [3], but the lack of an accurate antenna model has made the design interface between the circuit and the antenna challenging. To alleviate the modeling difficulty, $\lambda/2$ antennae are often employed in the design. However, for RF frequency bands such as 2.4-GHz, the length of a normal bondwire is much smaller than half wavelength. Although in theory reducing antenna size results in reduced performance, if the requirement of transmitting distance is only in short range and the bondwire antenna can be properly modeled, the approach of utilizing bondwire as an electrically small antenna is still feasible.

In order to model an integrated antenna, traditionally, they are required to be imported into three-dimension (3D) EM solver, run for S-parameter results, and finally fed into circuit solvers as an n-port network block. This flow has several inherent inconveniences. First, S-parameter results are represented in the frequency-domain. They cannot be directly applied to time-domain analysis. Although these results can be interpolated or extrapolated for time-domain analysis, this approximation is usually complicated and inaccurate at radio-frequency band because higher-order elements are required to model the transfer function over a large bandwidth. Second, since S-parameter data are port impedance-dependent, they

need a careful post-processing and have to be consistent with the interface impedances in the transmitter design testbench.

In this paper, a method which directly models bondwire antenna into an equivalent circuit is proposed. The generated equivalent circuit can be embedded into any other parts of linear or non-linear circuits, and simulated in circuit solvers to obtain either time- or frequency-domain results. In Section II, the background of PEEC theory is reviewed. Advantages of the proposed co-simulation method and a novel approach to obtain power transfer characteristic are proposed. In Section III, a 2.4-GHz CMOS transmitter IC prototype with a bondwire antenna is implemented and analyzed by the PEEC method. In Section IV, measurement results show that complex EM effects like bondwire antenna/circuits coexistence can be solved by this proposed method with high accuracy. Finally the benefits of the proposed methods in RFIC design are summarized.

II. DESIGN THEORY

A. PEEC Method

PEEC method was first proposed by Ruehli in 1974 [4] and originated from the electric field integral equation (EFIE)

$$\vec{E}^i(\vec{r}, t) = \frac{\vec{J}(\vec{r}, t)}{\sigma} + \frac{\partial \vec{A}(\vec{r}, t)}{\partial t} + \nabla \Phi(\vec{r}, t), \quad (1)$$

where \vec{E}^i is the incident electric field, \vec{J} is the current density in a conductor, \vec{A} is the magnetic vector potential, Φ is the scalar electric potential and σ is the electrical conductivity. In the consideration of a distributed system, PEEC model can be extended to retarded PEEC model which is formed by formulating the integral equation for the cell element which contains point \vec{r} according to

$$\begin{aligned} & \frac{1}{a_v \sigma} \int_v \vec{I}(\vec{r}, t) dv + \frac{\mu}{a_v} \int_v \int_{v'} G(\vec{r}, \vec{r}') \frac{\partial \vec{I}(\vec{r}', t_d)}{\partial t} dv' dv \\ & + \frac{\nabla}{\epsilon_0} \int_S \int_{S'} G(\vec{r}, \vec{r}') q(\vec{r}', t_d) dS' dS = 0, \end{aligned} \quad (2)$$

where v, v', S, S' are volume and surface cells, and t_d the delay time between two cell elements. $G(\vec{r}, \vec{r}')$ is derived by solving multilayer Green's function using computational method as in [5]. In (2), each term can be represented by the voltage or potential difference across the cell element. This transforms the sum of the electric fields in (1) into the Kirchhoff Voltage Law (KVL) over each volume and surface cell [4].

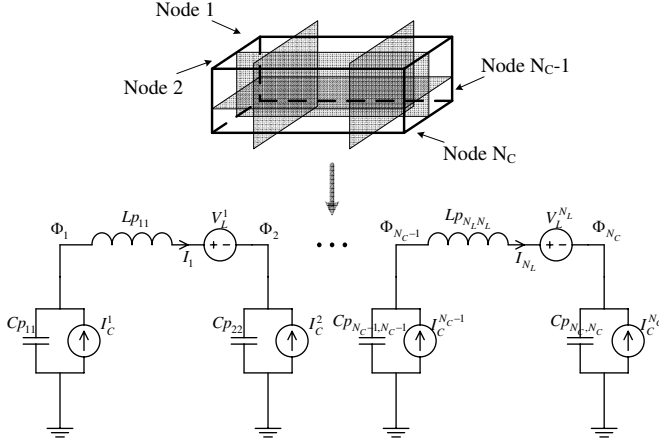


Fig. 1. Conductor equivalent circuits by PEEC method.

Fig. 1 details the retarded PEEC model of a cuboid conductor when discretized along 3D axes. The model consists of partial inductances Lp , partial capacitances Cp , retarded current controlled voltage sources V_L and retarded current controlled current sources I_C .

B. PEEC Method in EM/Circuit Co-Simulation

Traditionally, if a simulation system consists of both on-chip circuits and EM passive structures like an integrated antenna, the normal procedure is first to extract the S-parameters of the EM structure, followed by feeding the S-parameter data block into the circuit testbench to run simulations together with other circuits. For time-domain simulations, since S-parameter method is in the frequency-domain, S-parameter data are required across the whole operating spectrum or needed to be interpolated, and then converted into timing samples by the

Inverse Discrete Fourier Transform (IDFT) method which may cause inaccuracy if frequency samples are not fine enough. In contrast, the frequency-dependent effects of the retarded PEEC method is represented by a time delay in the values of the controlled sources in the EM structure's equivalent circuit model. This method can be regarded as providing a larger bandwidth model of the EM structure, which can directly produce time-domain results.

Furthermore, since S-parameter data are also dependent on source/loading impedances, any inconsistency between source/loading impedances in the S-parameter acquisition and the preceding/following impedances in the testbench will lower the accuracy of final simulation results. In the retarded PEEC method, the equivalent model is independent of the preceding and following impedances and can be seamlessly embedded into the circuit design testbench.

In Table I, equivalent circuits of a lump element EM structure and a distributed EM structure are compared. Lossless EM structures are considered here for simplicity. In the former case, if the EM structure is physically small compared to the wavelength, no frequency-dependent effects exist and it is a time-invariant, memoryless LC system with a pure negative value input impedance. All injected power into the EM structure are either reflected back or stored in its nearby region. Otherwise, if the size of the EM structure cannot be ignored, because of the existence of $e^{j\frac{2\pi}{\lambda}|r_1-r_2|}$ terms, input impedance shows both positive and negative parts. The positive part of this impedance is the radiation resistor and power dissipated on the radiation resistor equals to the total power radiated into space P_{rad} .

By transient (TRAN) or Periodic Steady-State (PSS) simulation, power level dissipated on the bondwire model can be simulated. This power level is the total input power P_i , which

TABLE I
FD AND TD REPRESENTATIONS OF AN EM STRUCTURE

System Cases	Equivalent Circuit	Controlled Sources in Freq. Domain	Controlled Sources in Time Domain
		$V_L^1 = j\omega Lp_{12} I_2^1 + \dots$ $I_C^1 = \frac{p_{12}}{p_{11}} I_2^1 + \dots$ $I_C^2 = \frac{p_{21}}{p_{22}} I_1^1 + \dots$	$V_L^1 = Lp_{12} \frac{\partial I_2^1(t)}{\partial t} + \dots$ $I_C^1 = \frac{p_{12}}{p_{11}} I_2^1(t) + \dots$ $I_C^2 = \frac{p_{21}}{p_{22}} I_1^1(t) + \dots$
		$V_L^1 = j\omega Lp_{12} I_2^1 e^{j\frac{2\pi}{\lambda} r_1-r_2 } + \dots$ $I_C^1 = \frac{p_{12}}{p_{11}} I_2^1 e^{j\frac{2\pi}{\lambda} r_1-r_2 } + \dots$ $I_C^2 = \frac{p_{21}}{p_{22}} I_1^1 e^{j\frac{2\pi}{\lambda} r_2-r_1 } + \dots$	$V_L^1 = Lp_{12} \frac{\partial I_2^1(t - \frac{ r_1-r_2 }{c})}{\partial t} + \dots$ $I_C^1 = \frac{p_{12}}{p_{11}} I_2^1(t - \frac{ r_1-r_2 }{c}) + \dots$ $I_C^2 = \frac{p_{21}}{p_{22}} I_1^1(t - \frac{ r_2-r_1 }{c}) + \dots$

includes both of the power consumed on conductance loss P_{loss} and radiated power P_{rad} . Antenna efficiency η_r , which is determined by the ratio of radiated power to the total power dissipated by the antenna, can be calculated [6] as

$$\eta_r = \frac{G_P}{D} = \frac{P_{rad}}{P_i} = \frac{R_{rad}}{R_{rad} + R_{loss}}, \quad (3)$$

where R_{rad} is the radiation resistance and R_{loss} the AC loss resistance considering of skin and proximity effects. As with any antenna design, the radiation resistance R_{rad} should be maximized while the loss resistance R_{loss} minimized. Feasible antennas require both reasonable radiation efficiency and the matching with preceding power amplification stage.

C. Novel Power Transfer Computation with PEEC Method

Once the characteristics of a TX antenna are obtained, Friis transmission equation can be used to estimate the power received by the RX antenna under idealized conditions [6]. However, this equation is only valid if the distance separated is much larger than the signal wavelength. For short distance, Friis equation is no longer accurate and power transmission characteristics have to be derived from Maxwell equations.

To overcome this inconvenience, our approach provides a unified procedure to calculate received power in any separated distance. After the acquisition of total input power P_i described previously, an arbitrary referenced far-field point is needed to pick up and the received power at this point $P_{far-field}$ can be obtained by the Friis equation

$$P_{far-field} = P_i \cdot \eta_r \cdot \left(\frac{\lambda}{4\pi R} \right)^2 \cdot G_r, \quad (4)$$

where R is the separated distance, λ is signal wavelength and G_r is the antenna gain of receiving antenna.

The received power at any separated distance point is determined by

$$\begin{aligned} P_{received} &= P_{far-field} \cdot \left(\frac{E_{received}}{E_{far-field}} \right)^2 \\ &= P_i \cdot \eta_r \cdot \left(\frac{\lambda}{4\pi R} \right)^2 \cdot G_r \cdot \left(\frac{E_{received}}{E_{far-field}} \right)^2, \end{aligned} \quad (5)$$

where $E_{received}$ and $E_{far-field}$ are the radiation electric field density at the observing and reference positions, respectively.

To calculate the radiation electric field density on an arbitrary point, our proposed approach is depicted in (6). The first step is also to solve the PEEC model for current of each volume cell. Then, we calculate a 1 by N_L matrix \tilde{L} between the observing point and each of the volume cells, using similar approach to the creation of L matrix in PEEC method. Finally, the electric field at this point is given by a dot multiplication of the \tilde{L} matrix and the volume cell current matrix. The advantage of this approach is that it applies to both near- and far-distance field calculations, without demanding Maxwell's equations in

any step.

$$\begin{aligned} \vec{E}(\vec{r}, t) &= \frac{\partial \vec{A}(\vec{r}, t)}{\partial t} \\ &= \mu \int_{v'} \frac{e^{j\omega t_d}}{|\vec{r} - \vec{r}'|} \frac{\partial \vec{J}(\vec{r}', t_d)}{\partial t} dv' \\ &= \frac{\mu}{a} \int_{v'} \frac{e^{j\omega t_d}}{|\vec{r} - \vec{r}'|} \frac{\partial \vec{I}(\vec{r}', t_d)}{\partial t} dv' \\ &= j\omega \begin{bmatrix} \tilde{L}_1 & \cdots & \tilde{L}_{N_L} \end{bmatrix} \begin{bmatrix} I_1 \\ \vdots \\ I_{N_L} \end{bmatrix}. \end{aligned} \quad (6)$$

With this approach, received power in any separated distance can be calculated. The results from this approach are consistent with Friis equation in long distance cases.

III. CIRCUIT DESIGN AND CO-SIMULATION

To demonstrate the validity of the method, a transmitter system with signals radiated by its bondwire has been designed. The transmitter IC consists of a 2.4-GHz oscillator and Class-D power amplifier, with block diagram as shown in Fig. 2.

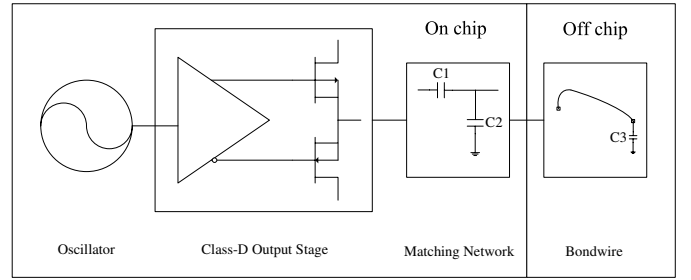


Fig. 2. IC prototype block diagram.

The shape of bondwire is consistent with Electronic Industries Association JEDEC standard [7] and was modeled according to retarded PEEC method. Total meshing quantity is 34 surfaces and 33 volumes in 3D axes with assumption that charges and currents distribute uniform in each surface and volume. Resulting partial element matrices were translated as a sub-circuit and simulated in the testbench with on-chip transmitter circuits simultaneously.

In order to radiate maximum power, the bondwire antenna is matched with preceding Class-D power amplification stage in Smith Chart by its equivalent circuit. As shown in Fig. 2, C1 and C2 are both implemented on chip for matching, and C3 is the only external component in the prototype and is used to tune the resonance of the transmitted signal. Radiation power varies significantly according to the value of external capacitor. However, with an accurate model of the bondwire, an optimum value of this capacitor can always be obtained.

In order to extend the application of this method to future general on-chip EM structures, a generalized platform was set up to conduct EM/circuit co-simulation based on an operating flow in Fig. 3. A MATLAB program has been developed to conduct retarded PEEC analysis and generate an equivalent

circuit netlist. This netlist can be fed into a circuit testbench to conduct simulations by Spectre simulator [8].

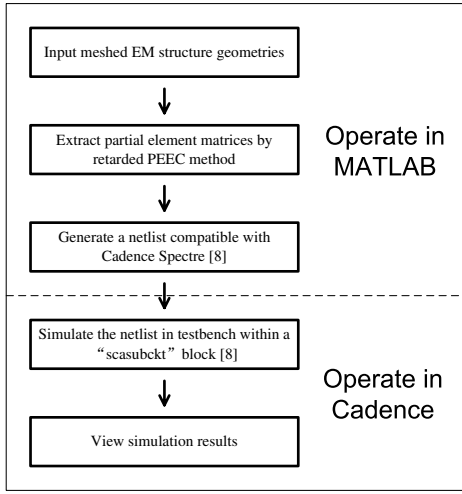


Fig. 3. EM/circuit co-simulation flow by retarded PEEC method.

IV. EXPERIMENTAL RESULTS

The transmitter IC was implemented in a $0.35\mu\text{m}$ standard CMOS technology. Chip area is $750\mu\text{m}\times 500\mu\text{m}$, and the bondwire as antenna has a length of 5mm and diameter of $25.4\mu\text{m}$. Fig. 4 shows its die photo.

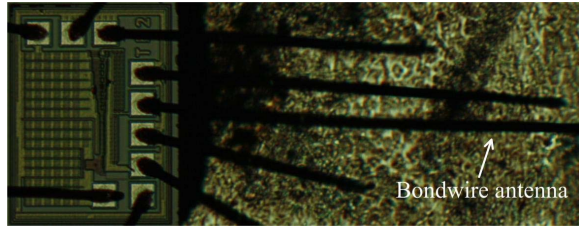


Fig. 4. Die photo of the transmitter IC.

The IC is mounted on FR4 PCB using a chip-on-board package. Two ends of the bondwire antenna are connected with chip pads and fed onto the external capacitor respectively. To build a communication link, a 2.4-GHz PCB patch antenna with 10dBi antenna gain connected to a spectrum analyzer is employed as the receiver. Wireless measurements were carried out in the anechoic chamber. Calibration procedure was firstly carried out with two standard antennae to find the path loss due to misalignment and polarization.

Based on PEEC modeling and co-simulation in the transmitter testbench, the total radiated power of the transmitter is -5.8dBm. The bondwire antenna gain is -15.5dBi, and its input impedance at 2.4-GHz is $2.89 + 120.77i$ ohms.

In Fig. 5, the received power derived by the proposed method and practical measurement are plotted. The red curve is obtained by retarded PEEC modeling of the bondwire antenna, then simulations of the generated netlist together with the other parts of transmitter circuits, and finally calculations

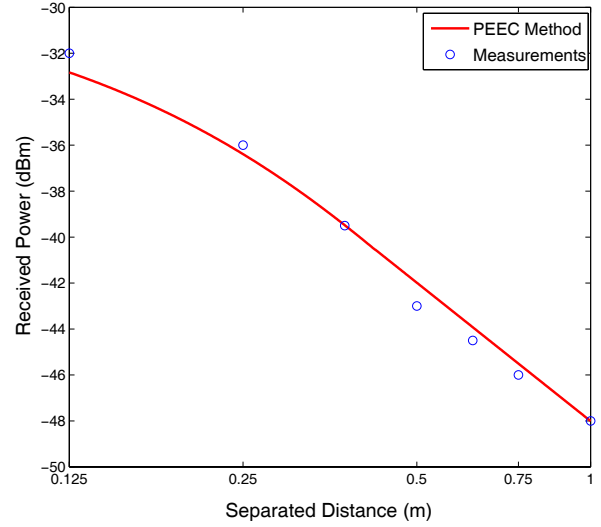


Fig. 5. Received power strengthes derived by the proposed method and practical measurement.

of the power transfer characteristics by the procedure mentioned in Section II. Blue circles are the measurement samples. Fig. 5 shows good agreements and verifies the accuracy of the proposed modeling method.

V. CONCLUSION

We have demonstrated, for the first time, an innovative application of the retarded PEEC method to model bondwire antenna for a transmitter IC prototype. The method provides accurate solutions and is convenient to deploy for RFIC and bondwire antenna co-simulation in both time and frequency domains. A generalized platform has been built to extend this modeling method as a promising EM/circuit co-simulation design tool.

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